RESEARCH DEPARTMENT

CRYSTAL-DIODE MODULATORS FOR THE FREQUENCY RANGE 10-1000 Mc/s

Report No. G-080

(1961/7)

THE BRITISH BROADCASTING CORPORATION
ENGINEERING DIVISION

RESEARCH DEPARTMENT

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SUMMARY

This report describes the development of low-power balanced diode modulators suitable for amplitude modulation by television or sound signals of a carrier in the frequency range 10–1000 Mc/s. From an original model, built in 1954, two later models have been developed having better performances in regard to the insertion loss at the carrier frequency, and uniformity of modulation response. The special transformers used are described.

1. INTRODUCTION

To facilitate the testing of television and sound broadcast receivers, a general-purpose diode modulator was developed in Research Department in 1954. The basic circuit subsequently found application in a complete modulator unit for N.T.S.C. colour television signals in Band I. Although the original diode modulator could be operated at frequencies up to about 1000 Mc/s, a satisfactory characteristic at frequencies above 400 Mc/s was somewhat difficult to obtain. Moreover, because of the high insertion loss, a small carrier leakage by unwanted paths could affect the modulator characteristic. An externally fitted coupling capacitor was required in series with the r.f. input of the first model when used with certain signal generators; in this case a disadvantage arose for television work because the uniformity of the modulation frequency response could be affected.

Further development was therefore carried out and two improved models were produced, one having greatly reduced insertion loss and good performance from about 30 to 250 Mc/s, and the other having a relatively low insertion loss and good performance over all broadcasting bands (Bands I to V) within the frequency range 40–1000 Mc/s.

2. PERFORMANCE REQUIREMENTS

A general-purpose modulator for v.h.f. or u.h.f. receiver testing should preferably be capable of modulating carriers in the frequency range 30–960 Mc/s, thereby covering all v.h.f. and u.h.f. broadcasting bands, as well as the usual i.f. frequencies of television receivers. The modulation frequency response should be substantially uniform over a frequency range extending from d.c. to about 10 Mc/s; experimental signals of different television line standards may then be generated. The modulator itself is inherently suitable only for double-sideband modulation; if required, vestigial sideband filters can be connected at the modulated signal output terminals. Good linearity should be maintained for depths of modulation

approaching 100%, with reasonable freedom from unwanted phase modulation. The specification in this respect varies according to the application, but may be exacting in the case of colour television signals.

In general, it is desirable for the modulator to operate with r.f. input and output terminals of 75 ohms impedance (unbalanced), and to have a low insertion loss at r.f.; an output level of up to 50 mV is sufficient for most purposes. The sensitivity with regard to the modulating voltage is not as important. Alternative modulation-input impedances of 600 ohms or 75 ohms are convenient to cover applications to audio frequency and television modulation respectively.

Since the type of modulator developed for this purpose was basically balanced, it was found possible, in the case of television, to adjust the biasing for either positive or negative modulation of the carrier, and to do this for either a positive-going or a negative-going modulation input. The facility of generating carrier-suppressed modulation is also available but is not required in testing broadcast receivers.

3. PRINCIPLE OF OPERATION

The modulator is basically a bridge circuit, incorporating diodes, which is balanced at the carrier frequency for certain conditions of bias on the diodes. The modulating signal is made to change the bias conditions and hence the diode impedance, so that a r.f. output is obtained through unbalance of the bridge.

It was found convenient to use a two-diode bridge. For this arrangement to work with good efficiency a balanced transformer is required, as shown in the basic circuit of Fig. 1(a). The two r.f. terminal pairs will be completely decoupled from one another if the slope resistances of the diodes, r_1 and r_2, are equal. In general, if a signal voltage V is applied to terminal 1, the open-circuit voltage at terminal 2 will be

$$\frac{r_1 - r_2}{r_1 + r_2} \cdot \frac{nV}{2},$$

where n is the turns ratio of the transformer.

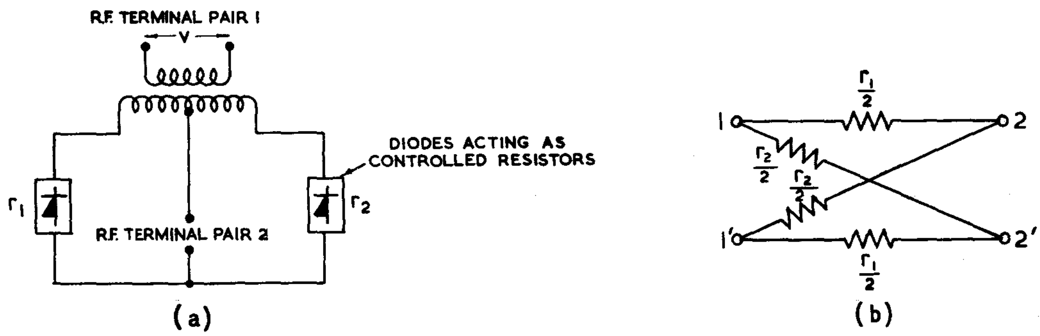


Fig. 1 - Principle of modulator circuit

(a) Basic circuit
(b) Equivalent circuit

If the device operates in the region, $r_1 \simeq r_2 = r$, say, the impedance seen at terminals 1 and 2 will be approximately $2r/n^2$ and $r/2$ respectively. The conditions for matching to terminals of resistance Z_0 are thus $r = 2Z_0$ and $n = 2$. Assuming operation near to this condition, i.e. $(r_1 - r_2) << (r_1 + r_2)$ and $r_1 + r_2 \approx 4Z_0$, the voltage at terminal 2 when loaded by Z_0 will be

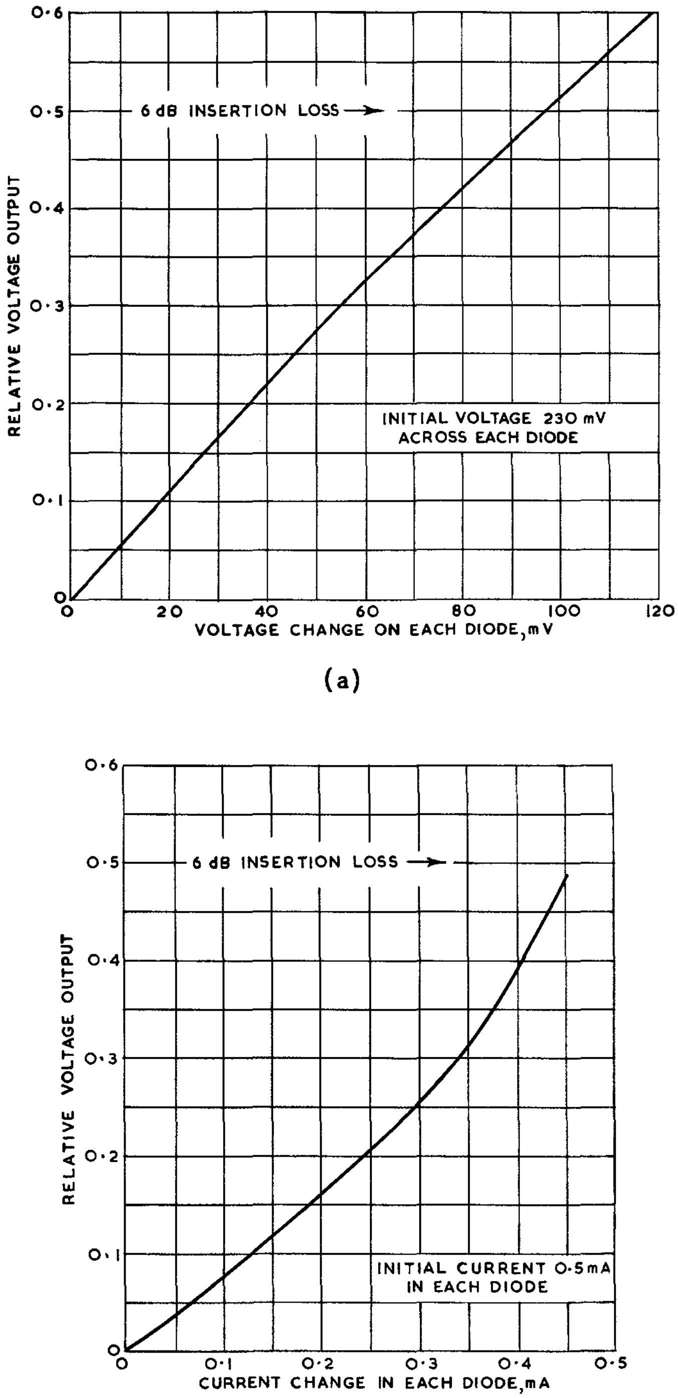
$$\frac{r_1 - r_2}{r_1 + r_2} \cdot \frac{V}{2} ;$$

the insertion loss of the circuit is therefore

$$20 \log_{10} 2\left(\frac{r_1 + r_2}{r_1 - r_2}\right) \text{dB.}$$

A more general analysis, based on the equivalent circuit shown in Fig. 1(b), shows that the network has an image impedance of $\frac{1}{2}\sqrt{r_1 r_2}$, and that if this is kept constant and equal to Z_0 the insertion loss is $17{\cdot}4 \tanh^{-1}\sqrt{r_1/r_2}$ dB. The same formula will, of course, apply equally to a r.f. signal applied to terminal 2 and taken from terminal 1.

Assuming terminal impedances of 75 ohms, a diode modulator of the idealized form considered would thus require diode resistances of the order of 150 ohms, and would also require the modulating signal to vary the resistances differentially in such a way that the geometric mean-value remains constant. This order of magnitude of impedance is readily achieved by a point-contact diode, with a small forward biasing current. Calculations of the insertion loss versus bias, based on the maker's characteristics for a type GEX66 diode, are given in Fig. 2.



(a)

(b)

Fig. 2 - Idealized modulator performance calculated from diode characteristics

(a) Voltage-controlled diodes
(b) Current-controlled diodes

These show that the linearity is somewhat better if the modulation controls the voltage across the diodes, rather than the current in the diodes. However, the non-linearity is of opposite type in the two cases, and supplying the modulation signal to each diode from a source impedance comparable with the diode impedance would be expected to give good linearity up to levels at which the modulator insertion loss is about 6 dB.

In a practical modulator it is convenient to adjust the diode bias to give linear modulation with a minimum of insertion loss at the extreme of the linear range of modulation; the mean diode impedance for the best compromise in this respect may differ by as much as a factor of two from the impedance for optimum matching, but this does not matter in the majority of applications. The exact conditions are affected by characteristics of individual diodes and by the circuit additions required to inject modulating and biasing currents, as described in detail later.

4. DESIGN OF BALANCED TRANSFORMER

An important assumption made in the previous theoretical discussion was that the transformer was ideal. A particular requirement for the efficient working in a two-diode modulator, which is not essential for an ordinary balanced-to-unbalanced matching transformer, is that the leakage reactance between the two halves of the split winding should be small, and that the lead-out inductances should be small. In other words, supposing the centre-tap to be earthed, there should be a low impedance to earth from the two ends of the winding if they are joined together.

Two different approaches were made to the design of a suitable transformer, the first based on an extension of low-frequency principles to high frequencies, using tape windings, and the second on an adaptation of a coaxial line "Pawsey stub" balanced-to-unbalanced transformer; in each case a ferrite core was employed.

4.1. Tape-Wound Transformer

The basis of construction is the use of a small ferrite core and thin copper tape for the windings, as illustrated in Fig. 3(a). At the time of design the smallest standard types of core available were in E and I forms, the outer arms of the E-pieces having $\frac{1}{8}$ in. by $\frac{1}{16}$ in. (3·2 mm by 1·6 mm) cross-section. For the transformer, L pieces were made by breaking off the outer arms of the E and grinding down to $\frac{1}{16}$ in. (1·6 mm) square cross-section. Two of these are used to form a closed ferrite core as shown in Fig. 3(b), with the windings on one of the longer arms.

The windings consist of thin copper tape, with shellac paper as insulation between the turns. The primary has four turns of $\frac{3}{32}$ in. (2·4 mm) wide tape. The secondary is made up with tape of half this width, so that the two halves, each of four turns, may be wound over the primary side-by-side. Moreover, by winding the halves in opposite directions, with a common connexion at the starting point representing the centre tap, very good symmetry between the halves of the split winding may be achieved. Fig. 3(c) shows cross-sections at the two levels. For clarity a reduced number of turns is shown on the diagram, and the circuit representation is given in Fig. 3(d). The inner winding is common to both diagrams of Fig. 3(c) but the halves of the outer winding appear separately in the two diagrams. An extra

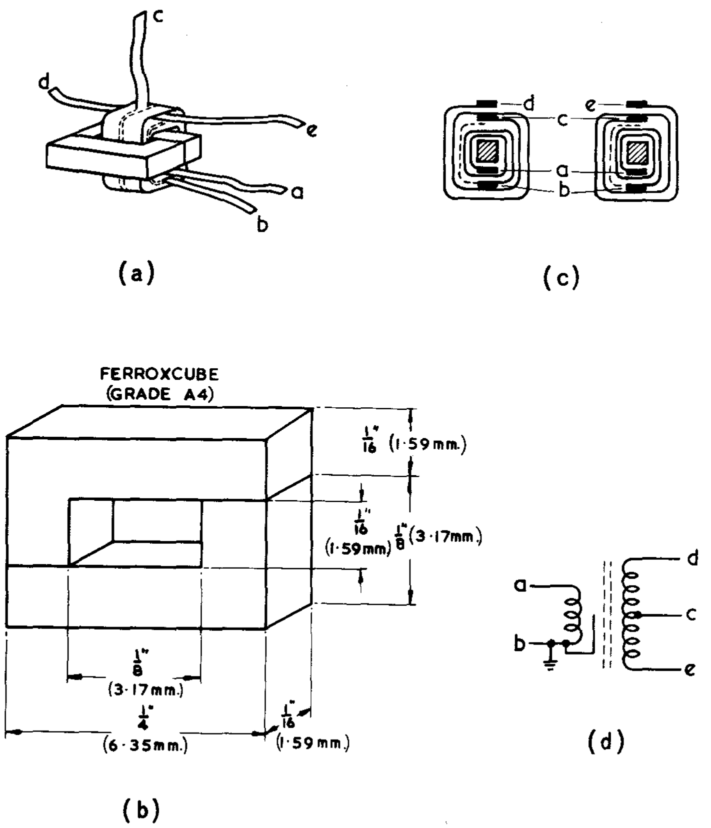


Fig. 3 - Tape-wound transformer

(a) General view
(b) Core dimensions
(c) Cross-sections at two levels
(d) Circuit representation

half turn is added to the inner winding beyond the point of connexion, b, (shown dotted). This is to reduce the net circulating current which flows via the inter-winding capacitance if a r.f. potential difference should occur between points b and c. A corresponding effect in the outer winding is compensated by the opposite action in the two halves.

It may be mentioned as a matter of interest that the extra half turn is not necessary in an ordinary tape-wound transformer if the outer winding is in the opposite sense to the inner; this is because currents induced by an inter-winding potential difference flow opposite ways round the core in the two windings. Alternatively, if the windings are in the same sense, an extra half turn is desirable both at the finish of the inner winding and at the start of the outer winding.

A four-turn winding on the core employed has an inductance (measured as a component in parallel with the loss conductance) of about 0·9μH up to 10 Mc/s. At higher frequencies the inductance falls, being about 0·3μH at 100 Mc/s. Above 100 Mc/s both the inductive susceptance and the parallel loss component remain

approximately constant, both components having a magnitude of about 5 mmho. The upper frequency limit of about 500 Mc/s is set mainly by the winding capacitance; both windings together are equivalent to a parallel capacitance of nearly 10 pF across the four turns. The corresponding susceptance rises to 30 mmho at 500 Mc/s, the nominal resistive load being 75 ohms or 13·3 mmho.

The measured performance of the transformer is discussed in Section 4.3.

4.2. Coaxial-Line Transformer

Miniature screened coaxial cable, with standard E and I ferrite sections, may be used to provide a transformer which is suitable for the present purpose, and which is somewhat easier to assemble than a tape-wound transformer. The principle of a 300/75-ohm balanced-to-unbalanced transformer of this type is best seen by first considering simpler types of transformer as shown in Fig. 4. Fig. 4(a) shows a well-known form of balanced-to-unbalanced transformer (sometimes called a Pawsey stub) employing quarter-wave coaxial lines or cables. A wide-band 75 ohms (unbalanced) to 75 ohms (balanced) equivalent transformer is shown in Fig. 4(b); the use of a ferrite core permits the use of short cables since the reactance of the balancing circuit is high over a wide band of frequencies. Fig. 4(c) shows a further adaptation giving a 2:1 turns ratio (i.e., a 4:1 impedance ratio) at low frequencies; as the result of using 150-ohm cable it acts as a 75 ohms (balanced) to 300 ohms (balanced) transformer

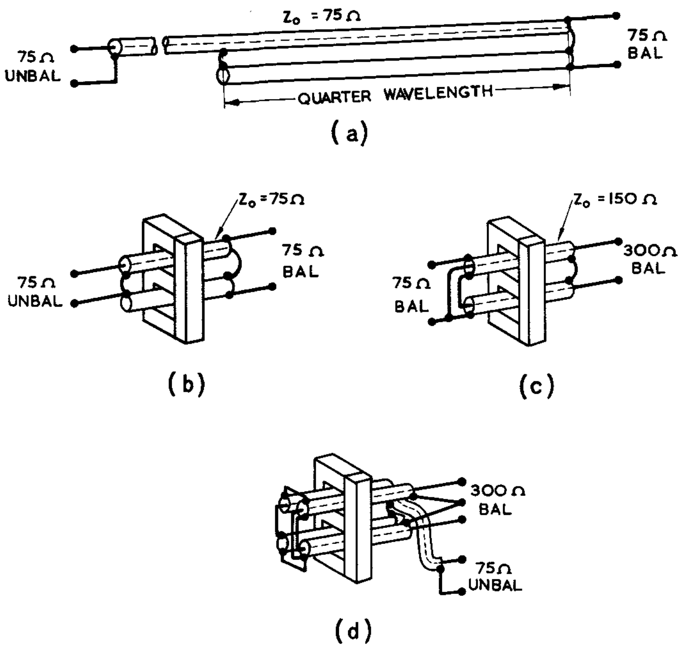


Fig. 4 - Coaxial-line transformers

(a) Quarter-wavelength balun
(b) Ferrite-cored balun
(c) Ferrite-cored balanced transformer
(d) Combined balun and transformer

over a wide range of frequencies. Transformers shown in Figs. 4(b) and (c) connected in cascade will give an arrangement for transforming from 75 ohms (unbalanced) to 300 ohms (balanced). In this case, since the same potential exists across one turn of each of the component transformers, they can with advantage share the same core. This leads to the "folded over" version shown in Fig. 4(d); it may be used only in a circuit which allows a common earth connexion for the primary and secondary.

In a practical transformer, it is possible to thread the cables through a series of transformer cores, the number depending on the mean frequency of the range to be covered. In the one employed for the modulator, a block of three E and I cores (Mullard type FX1052) was used; these cores have a maximum dimension of $\frac{1}{2}$ in. (1·3 cm). This arrangement gives a sufficiently high balancing-circuit inductance to give good performance down to about 30 Mc/s, and at the same time keeps the lengths of cable reasonably short. If the purpose of the transformer were merely to match between 75-ohm unbalanced and 300-ohm balanced terminals, and if ideal cable impedances of 75 ohms and 150 ohms could be used as indicated in Fig. 4, there would be no need to keep the lengths of cable short. However, if the secondary centre-tap is earthed, the "push-push" impedance to earth of the secondary (i.e., the impedance seen when the secondary terminals are joined together) is $j\frac{1}{2}Z_0 \tan\theta$, where $\theta = 2\pi l/\lambda_c$, l being the length of cable used on each side of the secondary winding, λ_c the wavelength in the cable and Z_0 the characteristic impedance. This push-push impedance should be low for the modulator application.

A compromise is necessary in regard to matching at the higher frequencies since only 50-ohm miniature cable (Callenders type T3202) was readily available; this was used for all the coaxial sections, and the low value of characteristic impedance helped to minimize the push-push impedance of the secondary, mentioned above. The parameter $\theta = 2\pi l/\lambda_c$ reaches about 30° at 900 Mc/s.

4.3. Tests on Transformers

The balance of the transformers was tested by measuring the insertion loss between a 75-ohm source and a 300-ohm load, using the arrangement indicated in Fig. 5. The two curves shown in Fig. 5 apply to the tape-wound and coaxial-line transformers already described. As a measure of the efficiency an additional measurement was made of the insertion loss between a 75-ohm source and load when the transformer is connected as shown in Fig. 6, using one half of the balanced winding.

In the case of the tape-wound transformer it was found important to make the measurements with the transformer mounted inside the modulator box. This is because the "live" ends of the balanced winding were in close proximity to the walls of the screened compartment. It was found that the properties are affected considerably by slight changes in the mounted position.

For good performance a balance of 30 dB is considered adequate, and it will be seen from Fig. 5 that, in this respect, the tape-wound transformer is satisfactory up to about 500 Mc/s, and the coaxial-line transformer up to about 1000 Mc/s. Regarding the one-side insertion loss shown in Fig. 6, the tape-wound transformer is generally superior at the lower frequencies, giving an insertion loss of only about 2 dB from 30 to 200 Mc/s. There is, however, a resonance effect which upsets the performance at higher frequencies, particularly above 450 Mc/s. The coaxial-line

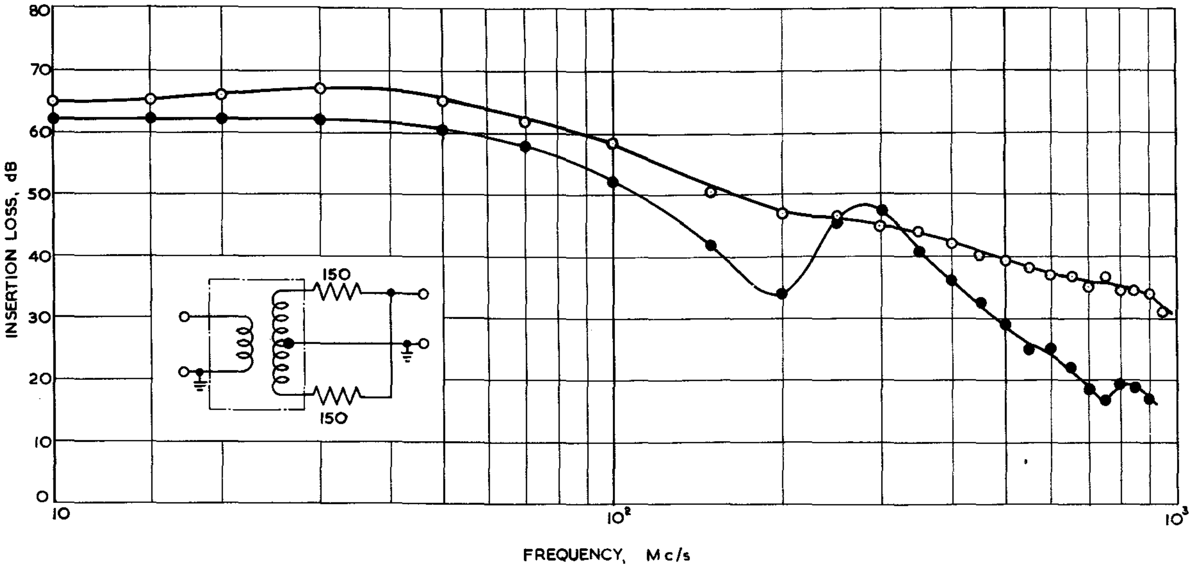


Fig. 5 - Transformer balance as a function of frequency

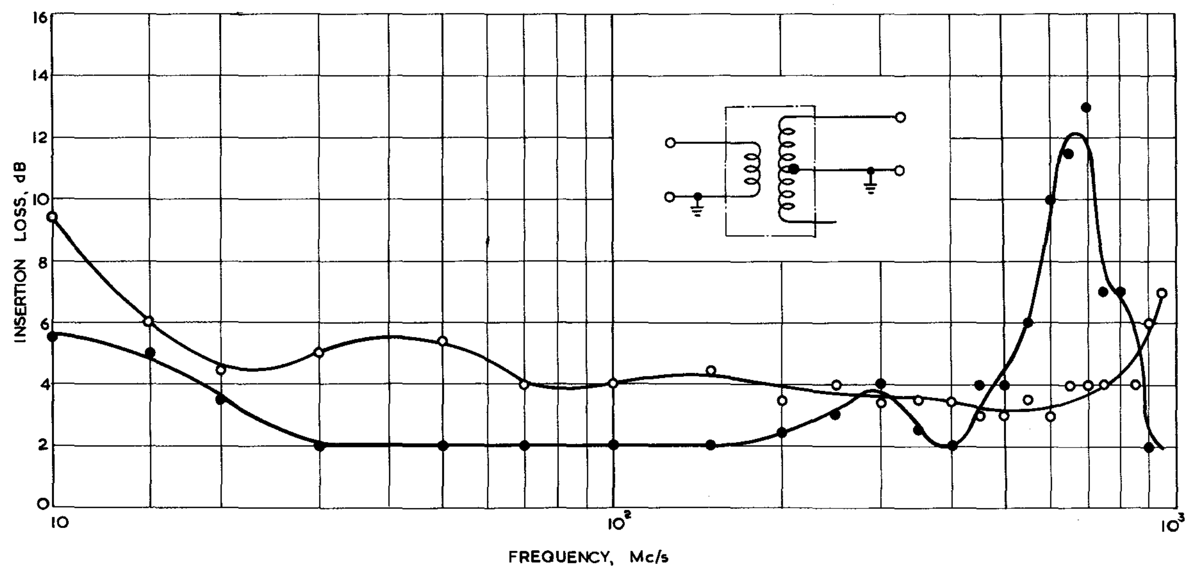


Fig. 6 - One-side insertion loss as a function of frequency

Tape-wound transformer

Coaxial-line transformer

transformer has a somewhat higher insertion loss at low frequencies but this remains in the region of 3 dB to 4 dB from 70 to 850 Mc/s.

These tests were made to examine the characteristics of the transformers which have been found to work satisfactorily in the modulator units to be described; the results do not necessarily indicate the best performance possible with either form of construction.

5. DESIGN OF MODULATOR

5.1. First Design (Type A)

The general-purpose modulator, as developed in 1954, is shown in Fig. 7; the circuit employed is shown in Fig. 8. The unit consists of two parts, a battery and meter cabinet, and the modulator unit proper mounted on it. The latter is contained within a small silver-plated brass box, provided with five separate screened compartments (shown on the circuit diagram of Fig. 8 by dotted lines). A detailed view of this box is given in Fig. 9. Four of the compartments form the main box while the fifth is brazed on top and carries the modulation input socket.

The two outer compartments of the main box house the bias filters and terminals for the bias supply leads from the battery box. The filters consist of a small choke (RB81) wound with copper tape and insulated with PTFE film, the core being

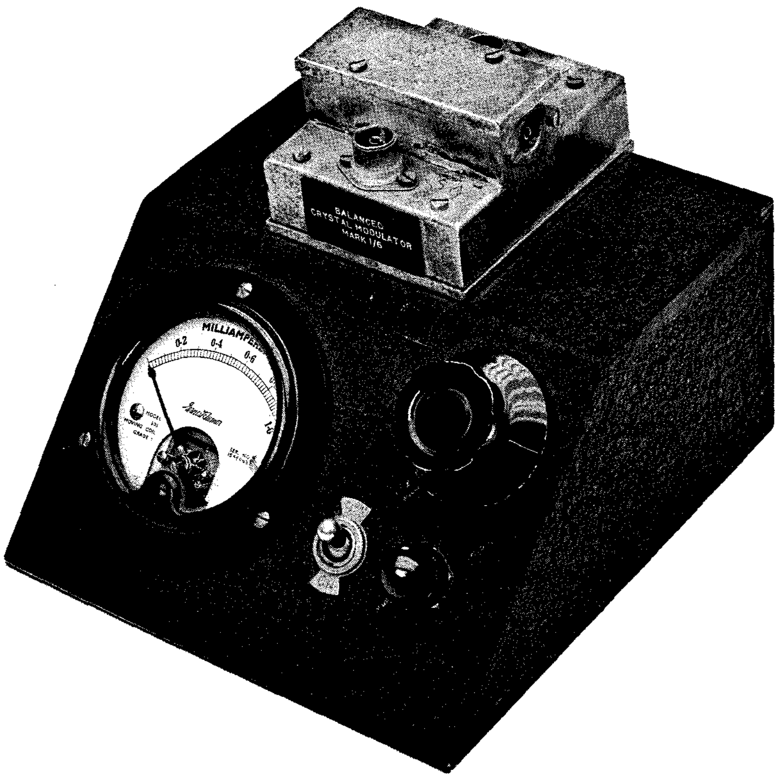


Fig. 7 - Complete modulator

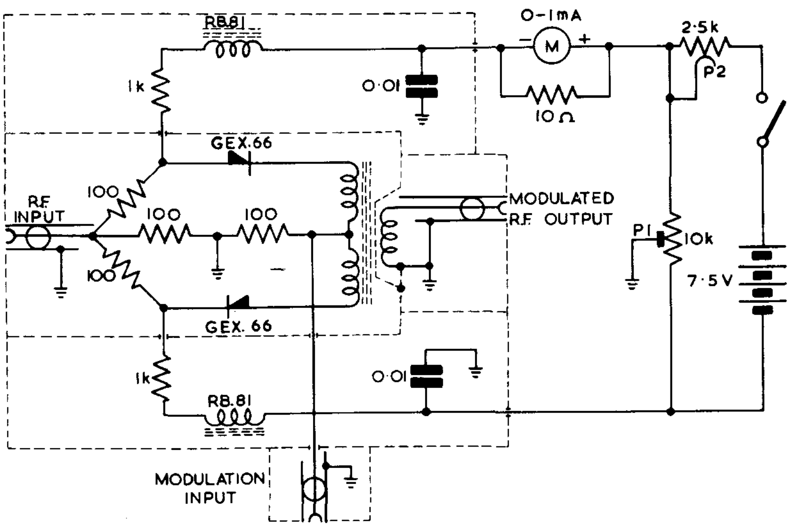


Fig. 8 - Circuit of Type A modulator

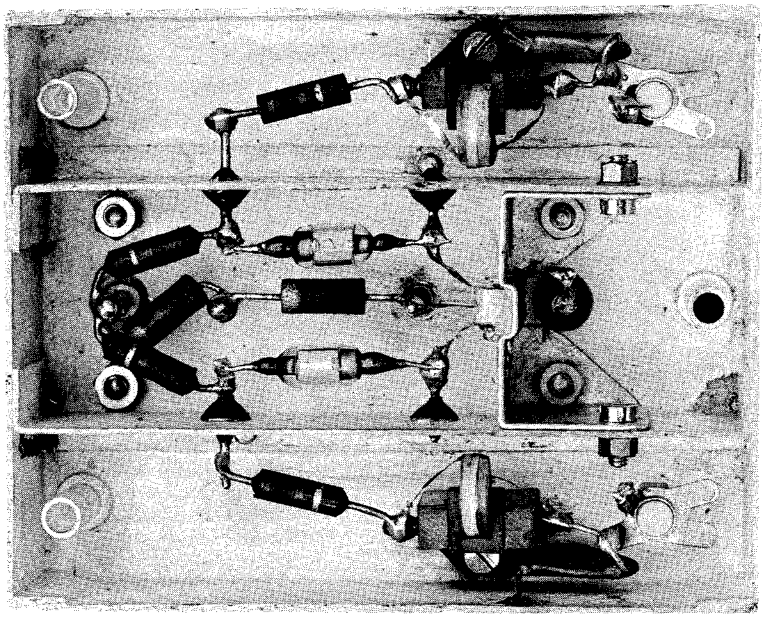


Fig. 9 - View of Type A modulator

made up from two L-shaped pieces of ferroxcube; at v.h.f. this choke represents an inductance of about 20 μH in shunt with a capacitance of 0·35 pF, together with a parallel resistance which has a value of 5000 ohms at 250 Mc/s. The other filter components are a 1000 ohm resistor and 0·01 μF disk capacitor, the object of the resistor being to prevent the modulation current, which flows through the capacitor via the choke, rising to a large value as the result of a resonance condition. The resistor also improves r.f. filtering at frequencies above 250 Mc/s. Bias is applied via a resistance network arranged so that the mean forward biasing current through the two crystals can be adjusted (by P2) as well as the differential bias (by P1).

The two central compartments contain the crystals (type GEX66), the resistance network, and the transformer, which is mounted on a screen dividing the two

compartments. The output connexions are contained in the second compartment; direct leakage between the r.f. input and output through stray capacitance or inductance is thereby prevented. The transformer is the tape-wound version described in Section 4.1.

With the circuit described the modulator is sensitive to any d.c. component superposed upon the modulation or r.f. inputs, so that precautions must be taken to ensure that they are free from undesired d.c. sources. In the case of the r.f. input, an adapter containing a series capacitor is provided. For audio frequency modulation, a transformer or "repeating coil" can be connected in the modulation chain immediately before the modulator. The modulator presents a load of about 75 ohms at its input sockets so that for modulation from a 600-ohm source a resistor of 500 to 550 ohms can be inserted in series with the modulation input socket; the input required for 100% modulation is then approximately zero level (1 mW into 600 ohms). The measured performance of the modulator is given in Section 7.

5.2. Second Design (Type B)

One disadvantage of the first design is that the insertion loss at the carrier frequency is high (about 25 dB). This means not only that good screening is needed, but also that it may be difficult to provide a modulated signal output of sufficiently high level. A serious loss arises from the use of the 100-ohm resistors in series with the diodes. Fig. 10 shows a re-arrangement of the circuit, using the same transformer, by which it is possible to approach much more closely the idealized r.f. circuit discussed in Section 3. This forms the basis of the Type B modulator.

The revised circuit includes d.c. decoupling from the r.f. input terminal, so that no special precautions are needed in respect of r.f. generators which might have a d.c. component at their output. However, it is assumed that the r.f.

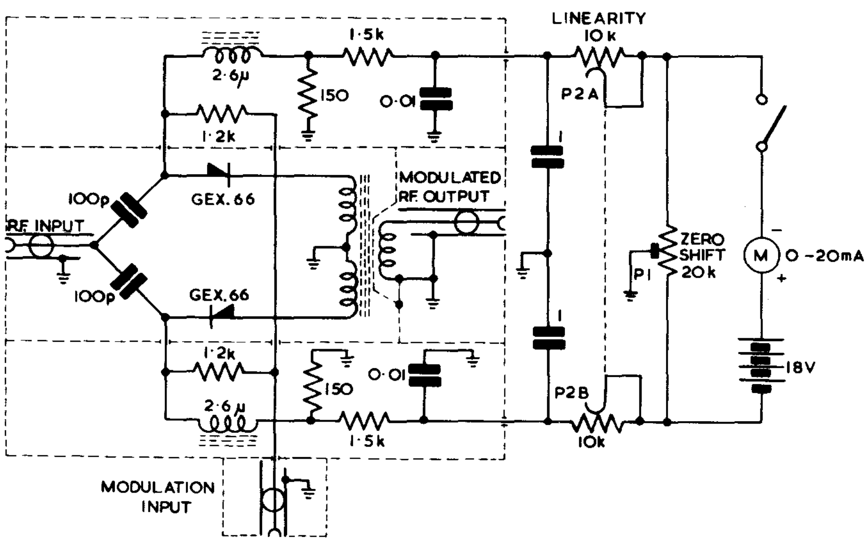


Fig. 10 - Circuit of Type B modulator

generator impedance remains close to 75 ohms (resistive) down to a frequency of about 1 Mc/s. The reason for this is that the r.f. generator, together with the 100 pF capacitors, the 2·6 μH chokes and the 150-ohm resistors shown in Fig. 10, should present a constant-impedance to the modulating signal, which is supplied via two 1·2 kilohm resistors. The modulation input impedance for the complete circuit is about 670 ohms; this is suitable for audio or television work. A shunt impedance of 82 ohms may be added at the input if a modulation input impedance close to 75 ohms is preferred.

The choke-inductors are of simple construction; they have eighteen turns of "Lewmex" covered wire wound in a single layer on a 0·15 in. (3·8 mm) diameter dust-iron former. Compared with the Type A modulator there is a greater degree of independence between the effects of the two bias controls; in particular, with the revised circuit, P1 is equivalent to adjustment of the d.c. level of the modulation input, regardless of the mean forward bias current, which is adjusted by the ganged control, P2A, P2B. As shown later, in Section 7, this modulator has very good performance and low insertion loss in Bands I, II and III, but does not maintain these properties above about 300 Mc/s.

5.3. Third Design (Type C)

With the modulator box slightly modified, and with a circuit similar in principle to that of Type B, it was found possible to use the coaxial-line transformer described in Section 4.2 in place of the tape-wound transformer. Initial experiments showed some rise in insertion loss at frequencies near 800 Mc/s. This was thought to be due to the capacitance of the lead-through points between the choke and diode compartment, and to deficiencies of the choke. The circuit was therefore modified slightly to the form shown in Fig. 11. This improved the performance at frequencies in the 500-900 Mc/s region, but at the expense of some increased circuit loss at frequencies which lie in the region of 350 Mc/s; however, since this frequency lies

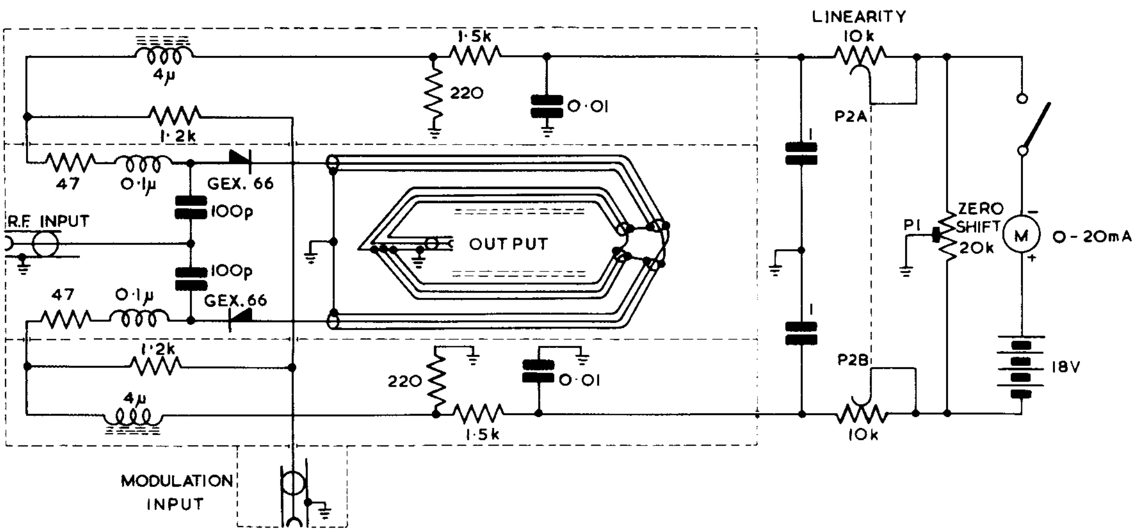


Fig. 11 - Circuit of Type C modulator

between Bands III and IV it was not considered important. The circuit change was an expedient to overcome difficulties associated with the particular components and the layout; with mechanical redesign the more straightforward circuit of Fig. 10 might well be satisfactory for frequencies up to about 1000 Mc/s.

6. ALIGNMENT AND OPERATION

The characteristics of point-contact germanium diodes vary considerably, and the initial setting up of the modulator involves examining the modulation characteristic. The most convenient method is to display the characteristic on an oscilloscope, using the arrangement shown in Fig. 12.

The signal generator is set to give an output not exceeding 100 mV. Larger inputs may affect the characteristic, as discussed in Section 7. As already stated, the Type A modulator is sensitive to any d.c. superposed upon the r.f. input or modulation input terminals, and precautions must therefore be taken to avoid this. The r.f. input may be conveniently protected by a series capacitor and the modulation input by a transformer. Modulators Types B and C are sensitive only to d.c. superposed upon the modulation input. The level of modulating signal to give 100% linear modulation is approximately "zero level" (i.e., 1 mW into 600 ohms), although higher levels may be used to explore the characteristic beyond the linear region. The Type A modulator should be adapted to present a modulation input impedance of about 600 ohms by the addition of a 560-ohm resistor in series.

The signal fed to the modulation input may for convenience be an audio frequency well within the range capable of being detected by the receiver. This signal is also applied to the X plates of the oscilloscope and should be adjusted to give the correct phasing on the oscilloscope display. The signal generator is square-

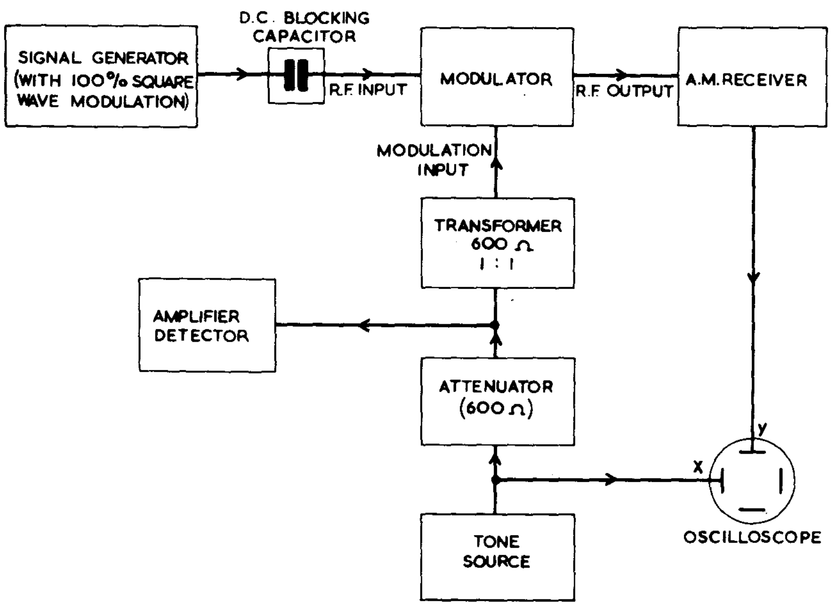


Fig. 12 - Block schematic of apparatus required to align a modulator

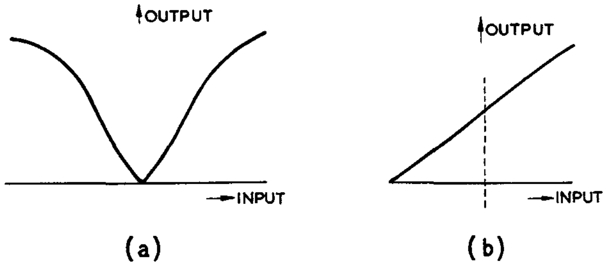


Fig. 13 - Modulation transfer characteristics

(a) Complete characteristic
(b) Portion used for linear modulation

wave modulated at a different audio frequency, which should also be small compared with the receiver bandwidth. The Y plate amplifiers of the oscilloscope should preferably be d.c.-connected to the diode detector of the receiver, which is tuned to the radio frequency of the signal generator.

Fig. 13 shows two of the display patterns which may be obtained by adjusting the input level of the modulating signal and the d.c. bias applied to the modulator. The base line of the display is provided by 100% square-wave modulation of the signal generator. If the modulating signal is made somewhat larger than "zero level", the characteristic displayed will appear as in Fig. 13(a), assuming that the initial biasing is approximately symmetrical.

It was found that, more particularly at higher frequencies, the minimum in this display could be blurred slightly. This indicates lack of balance in the stray circuit reactances when the diodes pass through the equal-conductance condition. It was found that selection of diodes could generally improve this feature. To obtain a very "clean" minimum at u.h.f. it was found necessary to introduce a small extra capacitance across either one diode, or across one half of the split winding of the transformer. This was done by means of a short length of insulated wire soldered to one terminal, and adjusted in proximity to the other terminal; the adjustment for a good minimum appeared to hold quite well over a wide frequency range.

For linear modulation a portion of the characteristic on one side of the minimum in Fig. 13(a) is used. The modulating signal is therefore reduced to a suitable level, the bias controls being adjusted to move the minimum to one side of the picture and also to achieve good linearity over the portion displayed, as shown in Fig. 13(b). The setting up is then complete for use in amplitude modulated "sound" carrier applications, and the level at the modulation input corresponds to 100% modulation. For reference purposes the insertion loss at the carrier frequency may be measured with the modulating signal removed. This gives the attenuation corresponding to the vertical deflection along the dotted line in Fig. 13(b). For "vision" carrier applications, the modulator loss for peak vision signal is 6 dB lower than the "sound" carrier loss measured in the manner described.

If the modulator is being set up for television applications a procedure similar to that described, using audio-frequency modulation, may be used. The horizontal deflection of a display like that of Fig. 13(a) may be calibrated in terms of the instantaneous voltage applied to the modulator. For symmetrical biasing this voltage would be zero at the horizontal mid-point. However, the modulator characteristic must in general be adjusted so that a linear portion lies over a practicable range of input voltages. That part of the characteristic which has a positive or negative slope may be used, depending on the required polarity of the modulation. While the d.c. component of the video signal must be properly preserved at the

modulator input, a large "standing" direct potential must be avoided, since the modulator bias controls are capable of making only small adjustments to the effective d.c. level.

7. PERFORMANCE

7.1. Insertion Loss

Fig. 14 shows the measured insertion loss of the three types of modulator as a function of frequency. The plotted results refer to the mean carrier amplitude. As explained in Section 6, for a television signal at peak amplitude the insertion loss would be 6 dB less than that shown by the curves. In each of the modulators there was a range of carrier frequencies over which it was no longer possible to achieve good linearity up to at least 97% modulation; adjustment was then made for the best compromise between the permissible depth of modulation and the linearity, the corresponding results for insertion loss being indicated by the broken line in Fig. 14.

It will be seen that, over the frequency range 40–250 Mc/s, the Type B modulator gives the lowest insertion loss, namely 12 dB referred to the mean carrier amplitude or 6 dB referred to the peak value. These values approach the theoretical results given in Section 3, which were based on an ideal transformer and a circuit using diodes conforming to the published characteristics.

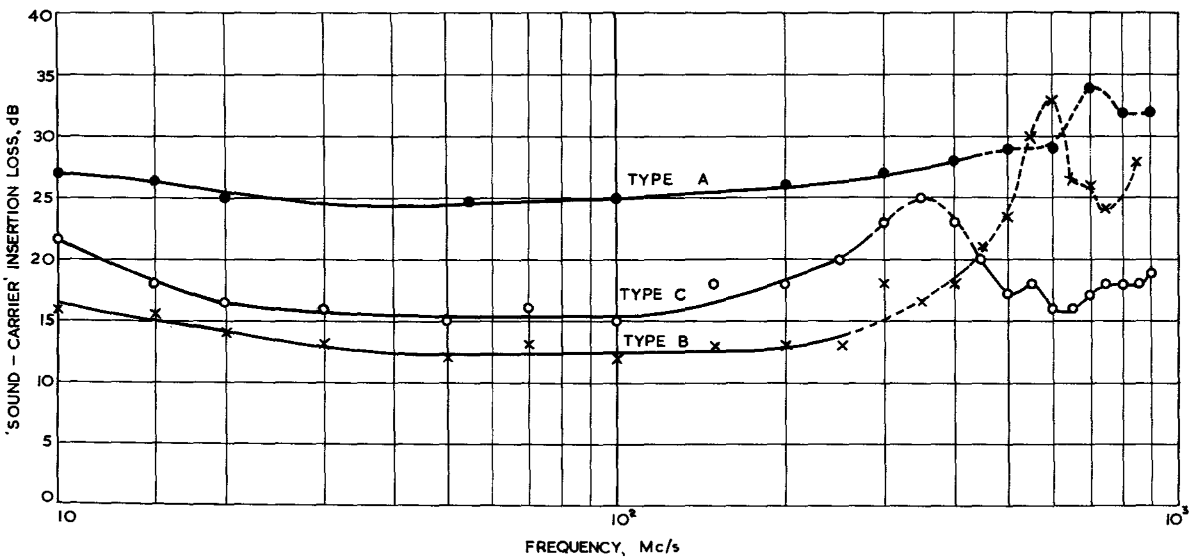


Fig. 14 - Insertion loss of modulator as a function of frequency

A broken line indicates a region over which there was difficulty in achieving linearity when close to 100% modulation

7.2. Modulation Frequency Response

The modulation frequency response of the modulators was measured over a wide range, with the primary purpose of finding the limitations in television applications. A method relying on the performance of a receiver and detector in regard to frequency response was not used since it is difficult to test the receiving equipment, particularly the detector circuit, without using a modulator of known performance. It was therefore decided to measure the amplitude of the sidebands directly, using a tunable receiver of good selectivity, the modulating signal being a sine wave of fixed amplitude, varied in frequency over the range 50 kc/s to 30 Mc/s. Figs. 15, 16 and 17 show results for the three types of modulator. In each case the amplitude of one of the sidebands relative to the carrier frequency is plotted. It was not permissible to rely on constancy of gain of the receiver at different frequencies, and amplitudes were therefore measured as changes in signal generator readings by making observations under the following four conditions:

(i) Receiver tuned to the carrier frequency output from modulator.

(ii) Signal generator connected directly to receiver, both tuned to the carrier frequency.

(iii) Receiver tuned to the sideband frequency from modulator.

(iv) Signal generator connected directly to the receiver, both tuned to the sideband frequency.

In cases (i) and (iii) the signal generator feeds the modulator at the carrier frequency at a standard fixed level. By adjusting the signal generator for a receiver output in (ii) equal to that in (i), and in (iv) equal to that in (iii), the required results may be obtained as the difference in signal generator output in cases (ii) and (iv). In view of the difficulties of a measurement of this type, errors of about 0·5 dB might be expected in individual points.

The rise in the frequency response for the Type A modulator between 200 kc/s and 1 Mc/s (Fig. 15) is caused by a 5000 pF capacitor connected in series with the r.f. input. This arrangement corresponded to a condition under which the modulator had been used in a number of applications but, in conjunction with the r.f. generator resistance, the series capacitor presents a frequency-dependent impedance which evidently affects the modulation response. Had this effect been fully appreciated at the beginning, a capacitance of lower value, or a r.f. generator free from a d.c. component in the output, and thus not requiring a capacitor, would have been employed.

The curves for the modulators Types B and C (Figs. 15 and 16 respectively) show that the frequency response is reasonably uniform. Since the revised circuit incorporates small capacitances at the r.f. input, the modulation frequency response remains uniform whether or not a large external capacitor is fitted in the r.f. feed, or is contained in the signal generator. The response is, however, still dependent on the r.f. source impedance remaining close to 75 ohms for modulation frequencies higher than about 1 Mc/s.

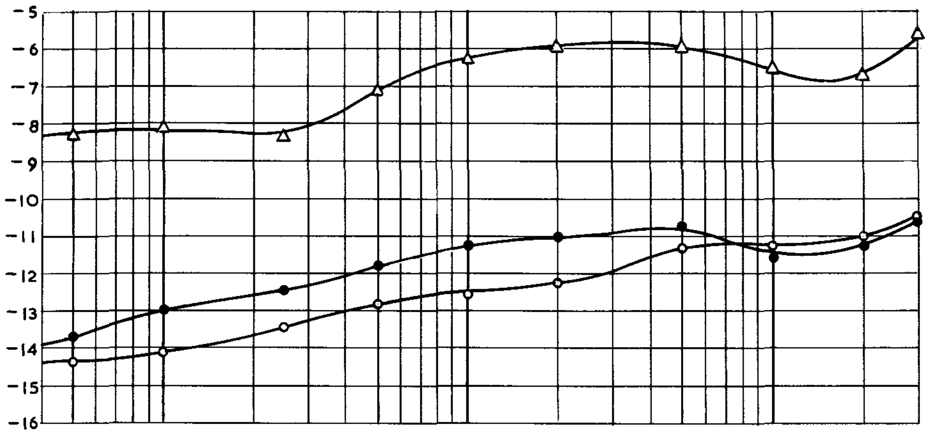


Fig. 15 - Modulation frequency response of Type A modulator

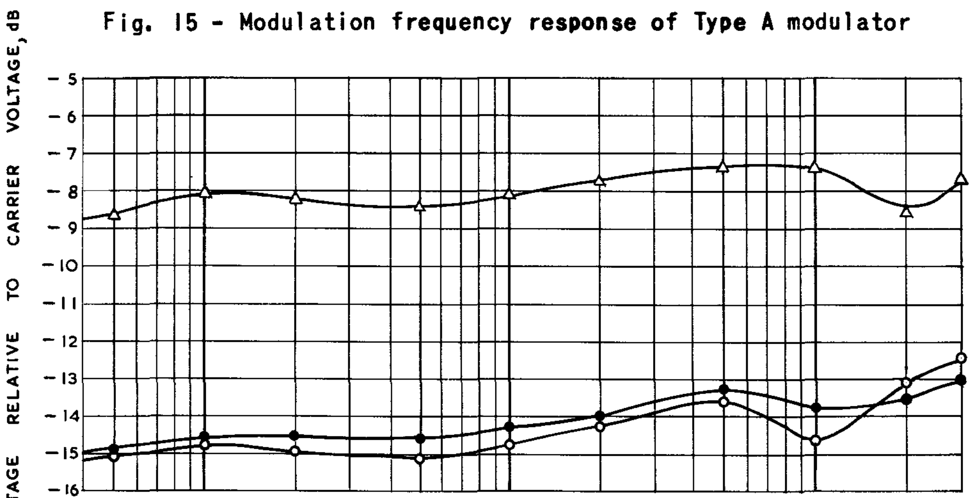


Fig. 16 - Modulation frequency response of Type B modulator

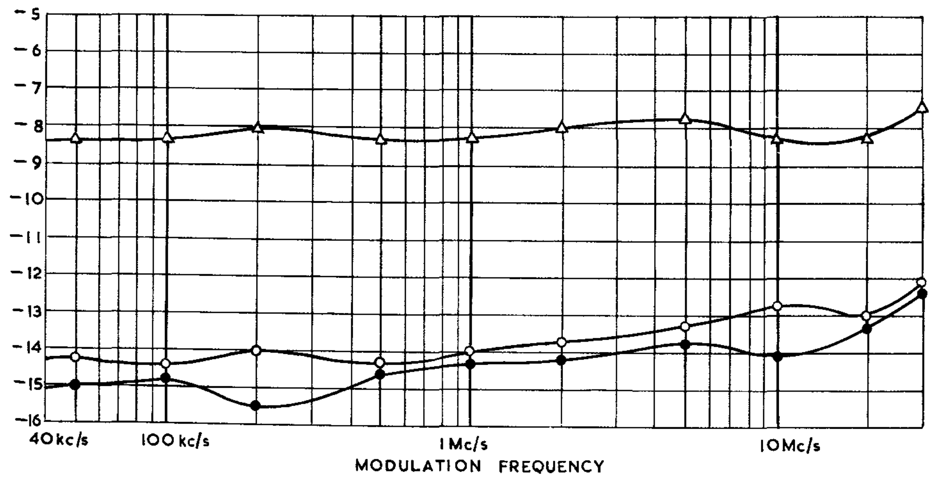


Fig. 17 - Modulation frequency response of Type C modulator

△ Upper sideband, 50 Mc/s carrier, approximately 90% modulation depth

● Lower sideband, 50 Mc/s carrier, approximately 40% modulation depth

○ Lower sideband, 150 Mc/s carrier, approximately 40% modulation depth

A difficulty also exists in connexion with the decoupling capacitors in the biasing circuit. For example, although no accurate experimental check has been made, a shallow step in modulation frequency response (under 1 dB) will occur at around 10 kc/s in the circuit of Fig. 8, or at around 100 c/s in the circuits of Figs. 10 and 11, because of the capacitors which decouple the biasing supplies.

In conclusion, it appears from the measurements at the greater depth of modulation, which are probably the more accurate, that the modulators Types B and C have a modulation frequency response which is uniform to ± 0·5 dB over the range 50 kc/s to 30 Mc/s. The experimental results at the smaller depth of modulation show a slightly greater fluctuation, but up to modulation frequencies of 5 Mc/s at least there is no real evidence of appreciable differences in the uniformity of frequency response from that at the greater depth. This result also gives some evidence of the symmetry of the sidebands, since different sidebands were used in the two series of measurements. It is not expected that the modulation response would be a function of carrier frequency, and this is confirmed (within experimental error) between the frequencies of 50 Mc/s and 150 Mc/s used for the measurements. More comprehensive tests in regard to this, and to the symmetry of the sidebands, were not considered necessary.

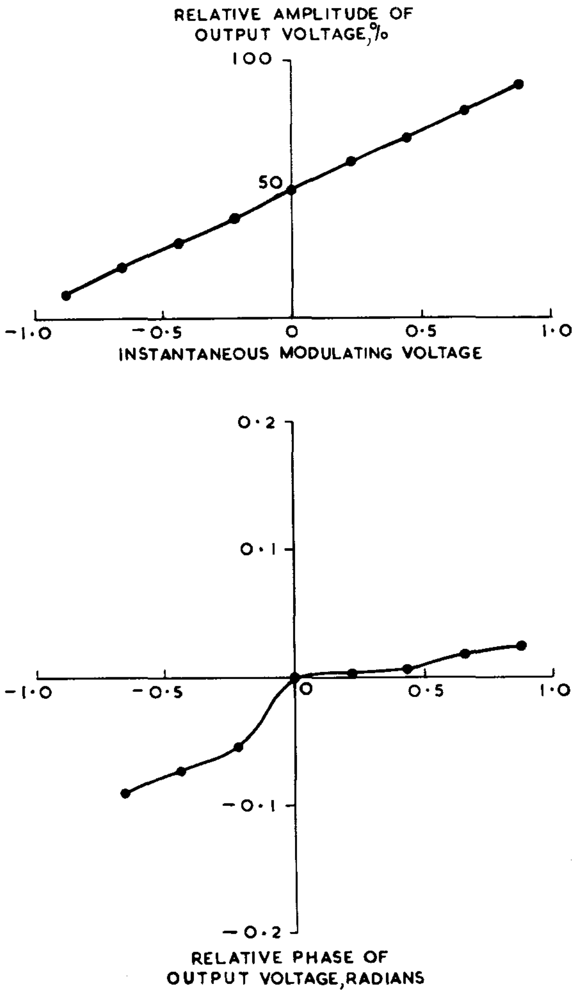


Fig. 18 - Amplitude and phase characteristics at 100 Mc/s for Type A modulator

7.3. Linearity and Unwanted Phase Modulation

The measurement of phase modulation and linearity was carried out in one operation under static conditions. A carrier signal was passed through the modulator and the relative complex amplitude of the output signal was measured with different d.c. potentials applied to the modulation terminals. The basic method of measurement employed has been fully described elsewhere.[1] The output from a signal generator is fed to a receiver by two parallel paths. One path includes the modulator and the other includes a General Radio admittance meter. The admittance meter can be arranged so that the attenuation of a signal passing through it can be adjusted in amplitude and phase. When adjusted for zero signal in the receiver, the modulator and admittance-meter outputs exactly cancel. The modulator characteristics may therefore be deduced from the readings on the meter.

An attenuator is placed in series with the modulator path, and is adjusted so that the minimum attenuation through this path is slightly greater than the minimum attenuation through the admittance meter path, the modulator having been set up to operate over a typical linear range. Direct-current potentials between the positive and negative limits of modulation are applied to the modulator from a source impedance of about 600 ohms, and the relative amplitude and phase of the r.f. output of the modulator is measured. Measurements were made for a carrier frequency of 100 Mc/s on the Type A modulator; the relative amplitude and phase are shown plotted against modulating voltage in Fig. 18. The phase has been normalized to be zero for "zero" modulating voltage, i.e., that corresponding to the mean carrier level; this level is 50% of the maximum amplitude in the range of linear operation. Similar measurements were also made at 100 Mc/s, 350 Mc/s and 800 Mc/s for the Type C modulator. It was found that the amplitude characteristics were reasonably linear; Fig. 19 therefore gives only the phase variations, plotted against the relative output amplitude. It will be seen that phase changes of the order of 0·1 radians at 100 Mc/s or 0·2 radians at 800 Mc/s occur between output levels of 30% and 90% of the peak output.

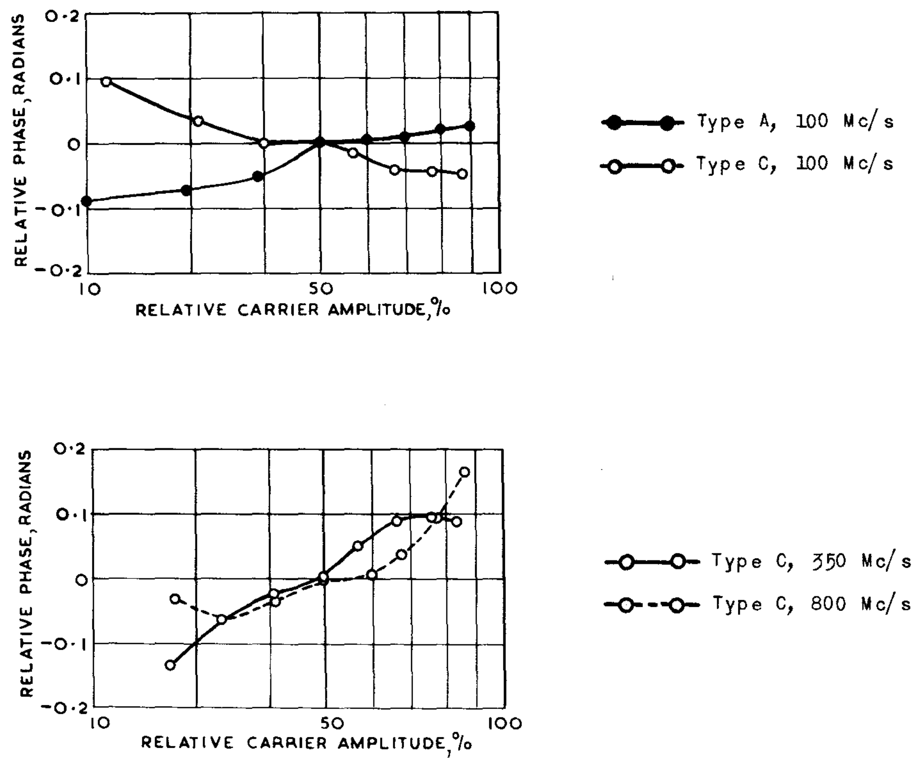


Fig. 19 - Modulator phase characteristics

7.4. Output Impedance

Measurements were made of the output admittance of a Type A and a Type B modulator when set up for normal linear modulating conditions, but with the input signals removed. The input signal generators were replaced by load resistors of the appropriate impedance, and the admittance looking into the output socket was measured

over a range of frequencies. The measured admittance was inverted, and expressed as an impedance, normalized to 75 ohms, referred to the rear of the output socket. The results are shown plotted on an impedance circle diagram in Fig. 20.

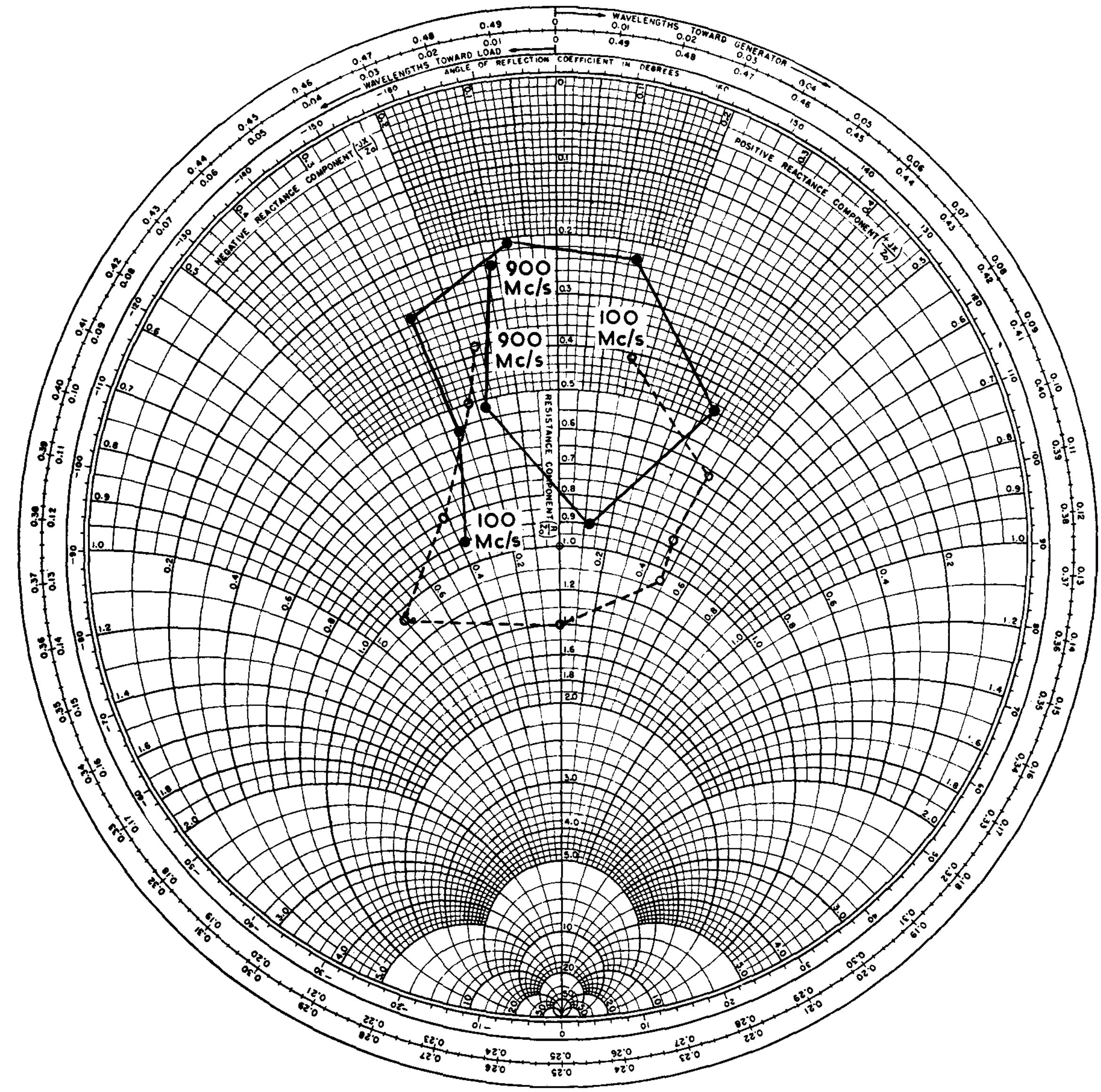


Fig. 20 - Modulator output impedance
(Impedances normalized to 75 ohms)

Type A modulator

Type C modulator

7.5. Maximum Output

Observations of the modulator characteristics, using the method described in Section 6, show that no appreciable change occurs for inputs up to 100 mV, and that the insertion loss remains constant. The Type B modulator at 65 Mc/s showed

a rise in insertion loss for an input of 200 mV, but the modulation linearity was only slightly affected. Under this condition the main carrier output level was 45 mV. Higher inputs caused saturation, which affected the characteristic. At the same time the mean-carrier output level tended to remain constant at about 50 mV. A similar behaviour was observed on a Type C modulator operating at 600 Mc/s.

It thus appears that the maximum output for the modulators operating normally corresponds to a mean modulated-carrier level of 30 mV, and that only a slight increase in output (with possible deterioration in linearity) can be obtained by increasing the input carrier level.

8. DISCUSSION AND CONCLUSIONS

It has been shown that, with care in the design of the necessary r.f. transformer, a two-diode modulator is a practical proposition for the generation of double-sideband amplitude modulated signals at low level in the v.h.f. and u.h.f. broadcast bands. Applications include amplitude modulation of a f.m. signal at audio frequency, for measurement of the a.m. suppression of f.m. receivers, and the generation of modulated "sound" and "vision" signals for testing television receivers.

A limitation that requires further discussion is the unwanted phase shift as a function of amplitude ("differential phase"), and its significance in the case of a colour television signal of the N.T.S.C. type. In this application it is the phase of the modulation at a subcarrier frequency of a few megacycles per second that is important, and this is governed by the phase relationship between the carrier-frequency component and one sideband component. We will first consider the phase errors in absolute terms, and then consider how a change in the amplitude of the vision carrier may affect the relevant phase relationship.

If the characteristic of the modulator is such as to produce a small amount of phase modulation *in step* with the amplitude modulation, the two sidebands remain equal in amplitude but differ in phase from those generated in a purely amplitude-modulated signal. A peak phase modulation of ± 0·1 radian accompanying a ± 10% amplitude change, for example, would produce sidebands differing by 45° in phase from the ideal. If the r.f. phase of the modulator output varies linearly with the logarithm of the amplitude, the sideband phase-shift is unaffected by variations of carrier level caused by picture modulation, and is also substantially independent of the depth of amplitude modulation. For this case, the phase shift would not affect colour fidelity of a N.T.S.C. picture because the "reference burst" by which the phase zero is established would be subject to the same phase shift. On the other hand, a non-linear relationship between the r.f. phase and the logarithm of the amplitude leads to changes in phase of the sidebands, and hence changes in the subcarrier phase after single-sideband demodulation, as the mean carrier amplitude or picture brightness changes. For convenience, Fig. 19 has been plotted to show the variations of phase as a function of the logarithm of the amplitude, so that the phase errors may be more easily assessed. Bearing in mind that the finer irregularities are probably due to experimental errors, the Type A and Type C modulators are shown to have reasonably linear phase characteristics at 100 Mc/s, and the resulting phase changes are likely to be well under 10°. However, the Type C modulator shows more serious errors at higher frequencies and, from the changes in slope of the smoothed charac-

teristic, it appears that the "differential phase" may reach 30° for a carrier frequency of 800 Mc/s.

In general, it may be concluded that the performance of the modulator is satisfactory for generating colour television signals at Band I frequencies (which are below 100 Mc/s), but for the higher-frequency bands (Bands III, IV and V) further work would be necessary to see whether the modulator could be aligned to obtain a satisfactory performance in regard to the change in phase with amplitude of the colour subcarrier. For this purpose it might be useful to add a control to adjust the reactive balance of the modulator circuit.

9. ACKNOWLEDGEMENTS

The modulators described owe much to the ideas of several colleagues of the authors, notably to Mr. G.D. Monteath who first pointed out the advantages of a balanced diode modulator, and to Mr. G.F. Newell who was responsible for producing the first model. Mr. C.G. Mayo and Mr. R.H. Minns gave advice on the design of the transformers.

10. REFERENCE

1. Monteath, G.D., Whythe, D.J. and Hughes, K.W.T., "A Method of Amplitude and Phase Measurement in the V.H.F./U.H.F. Band", Proc.I.E.E., Vol. 107, Part B., No. 32, pp. 150-154, March 1960.

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